Wireless Personal Communications **11**: 63–77, 1999.

Maximizing the SDMA Mobile Radio Capacity Increase by DOA Sensitive Channel Allocation

CHRISTOF FARSAKH and JOSEF A. NOSSEK
Institute for Network Theory and Circuit Design, Technical University of Munich, Germany

Abstract. In a cellular mobile radio system an SDMA[1] component can be implemented for the reuse of radio channels physically incorporated by time, frequency or code slots. Since SDMA is based on the spatial separation of different users operating in the same channel, a DOA[2] sensitive channel allocation scheme is essential for maximizing system capacity. In this paper we present the *Eigenvector method*, a computationally efficient algorithm to do this job. We also present simulation results on the *Eigenvector method* operating in a typical urban mobile radio cell. The simulated blocking probabilities are then used to predict the capacity increase which can be expected after adding an SDMA component to a conventional mobile radio system.

Keywords: wireless communication, smart antennas, SDMA, channel allocation.

1. Introduction

The basic idea of SDMA is the reuse of a radio channel – incorporated by an FDMA, TDMA or CDMA[3] slot 0- by $K > 1$ different users in the same mobile radio cell.

On the uplink, the spatial separation of K signals can be done by exploiting the information supplied through an $M \geq K$ element antenna array at the base. On the premise of roughly synchronized user bursts, the data sampled at the array can be used to jointly estimate the fast fading channel impulse responses relevant for all K user-specific uplink channels. These channel estimates can then be fed to a linear [1] [2] or a non-linear [3] [4] data detector yielding estimates for the symbols transmitted by each user.

Alternatively, DOA based spatial filtering techniques can be employed to separate the K signals transmitted on the uplink. K single user data detectors operating on the separated signals can then produce estimates of the data symbols transmitted by each user [5].

On the downlink, the spatial separation of the users is more difficult, since the mobiles cannot necessarily be expected to do joint detection and adaptive interference cancellation:

- The mobile terminals must be cheap, small and light. Preferably, the mobile station will not be equipped with an antenna array and sophisticated joint detection signal processing hardware.
- In general, the propagation delay of the mobile radio channel varies from user to user. On the uplink, user-specific transmission timing can easily compensate for these delays. On the downlink, the K signals transmitted by the base station cannot be timed in a way that each user receives his signal synchronized with the interference from the signals bound

[1] Space Division Multiple Access.
[2] Direction of arrival.
[3] Frequency, Time and Code Division Multiple Access.

for the $K-1$ co-users receiving in the same channel. Therefore, the mobile has little chance to effectively execute interference cancellation.

For these reasons, the appropriate way to do spatial separation on the downlink is to prevent interference for all K users by beamforming at the base station antenna array. The beamforming weights are supposed to be controlled by the spatial channel parameters estimated by means of the K user signals received on the uplink.

In many mobile radio systems the channel impulse responses estimated on the uplink cannot be directly reused as beamformer inputs due to the frequency and/or time gap between the uplink channel and the downlink channel [6] [7] [8]. In fact, channel reciprocity is only given for the DOAs and the corresponding medium term averaged path attenuations. Therefore, the adaptive control of the beamforming weights has to be based on the uplink channel estimates averaged over the fast fading.

The medium term downlink channel of each user $k = 1 \cdots K$ can be efficiently described by means of the $M \times M$ spatial covariance matrix

$$C_k = \sum_{q=1}^{Q_k} A_{kq}^2 a_{kq} a_{kq}^H. \tag{1}$$

The number of propagation paths between the location of the base station and the location of the user k is denoted by Q_k. Each path $q = 1 \cdots Q_k$ is described by its average squared transmission factor A_{kq} and the array steering vector a_{kq} incorporating its DOA, represented by an azimuth-elevation pair (ψ_{kq}, θ_{kq}). Efficient algorithms named Unitary ESPRIT and 2D Unitary ESPRIT to estimate these parameters in real time were presented in [9] and [10].

The spatial covariance matrices $C_1 \cdots C_K$ do not contain any information about the fast fading relevant for downlink transmission. Therefore, the average SNIR [4] necessary for each downlink receiver will be higher than the average SNIR the base station antenna array can cope with during uplink reception.

This implies that in contrast to most conventional mobile radio systems, the SDMA downlink, not the SDMA uplink, will be the critical link. Hence, the SDMA capacity increase gained by reusing the resources within a cell has to be evaluated according to the average number of users which can spatially separated by beamforming on the SDMA downlink.

2. Spatial Separability

On the downlink, the spatial separation of K different users can only be managed in a robust way, if the DOAs of all users in one channel are not too close to each other. Otherwise, downlink transmission will face severe problems, which can be illustrated by the following example: $K = 2$ users, each characterized by a single propagation path with identical attenuations, have to be separated by means of an $M = 4$ element uniform linear antenna array (ULA). In case (a) the azimuths ψ_1 , ψ_2 of the two corresponding DOAs are quite different from each other ($\psi_1 = +30^0$, $\psi_2 = -30^0$), whereas in case (b) they are almost identical ($\psi_1 = +3^0$, $\psi_2 = -3^0$).

The beampatterns created by two optimized weight vectors w_1 and w_2 yielding an average SNIR of 10 dB for both users are depicted in Figure 1. In case (b) the steep slopes of the patterns at the DOAs of the users ($\pm 3^0$) make two problems obvious:

[4] Signal-to-noise-and-interference ratio.

1. The performance of the downlink beamformer is extremely sensitive to DOA estimation errors.
2. The DOAs of the users are far away from the maxima of the corresponding beampatterns. This waste of electromagnetic power will result in unnecessary CCI in neighboring cells.

In this context, it makes sense to call the case (a) a "spatially well separable" scenario and the case (b) a "spatially badly separable" scenario.

The problems occuring in scenarios like (b) can be avoided by combining SDMA with at least one different multiple access scheme like FDMA, TDMA or CDMA which can supply the system with a number L of separate channels. A DOA sensitive channel allocation algorithm can then assign spatially badly separable users to different channels.

3. DOA Sensitive Downlink Channel Allocation

Let us assume that prior to a new user's channel request there are $K^{(l)}$ users operating in each channel $l = 1 \cdots L$. The numbers $K^{(l)}$ are not necessarily equal. In order to avoid trivial solutions there are no vacant channels assumed ($K^{(l)} \neq 0$). Altogether $J - 1 = K^{(1)} + \cdots + K^{(L)}$ users are on air.

After adding a new user to the system, L^J new user-channel combinations will be possible, provided there are no restrictions concerning reallocations of the $J - 1$ active users. In this case, the number of combinations to be checked will be prohibitively high even in small systems (e.g. $7 \cdot 10^{11}$ for $L = 7$ and $K = 14$). Another reason, why we will not consider any user reallocations during the allocation procedure, is the additional signalling traffic caused by intracell handovers.

Therefore, we will solve the SDMA channel allocation problem by a two-step procedure: First, finding the channel $l_{\text{opt}} \in \{1 \cdots L\}$ with best spatial separability for the new user; second, evaluating the spatial separability in that channel and then decide whether the new user will be allocated to the chosen channel or whether his channel request will be rejected.

The quality of any SDMA channel allocation scheme has to be evaluated according to how far the following goals can be achieved:

1. Maximizing system capacity by maximizing the average number of users which can be accommodated in L channels.
2. Ensure high probability for robust downlink communication by maximizing robustness of the beamformer against parameter estimation errors.
3. Minimizing CCI in neighbouring cells by minimizing the average RF power emitted by the base.

A computationally efficient algorithm doing this job is given by the *Eigenvector Method* (named "Quick SB algorithm" in [11]). It is based on the considerations presented in the following sections.

3.1. The SDMA Downlink Beamforming Problem

Let us assume we want to supply one user, characterized by the receiver noise N (composed of thermal receiver noise and CCI from neighboring cells) and the spatial covariance matrix C, with a given signal-to-noise-and-interference ratio *SNIR* by applying the complex weight vector $\boldsymbol{w}$ at the M element base station antenna array. The weight vector $\boldsymbol{w}$ will be chosen in a way that it minimizes the downlink transmit power P that has to be emitted at the base.

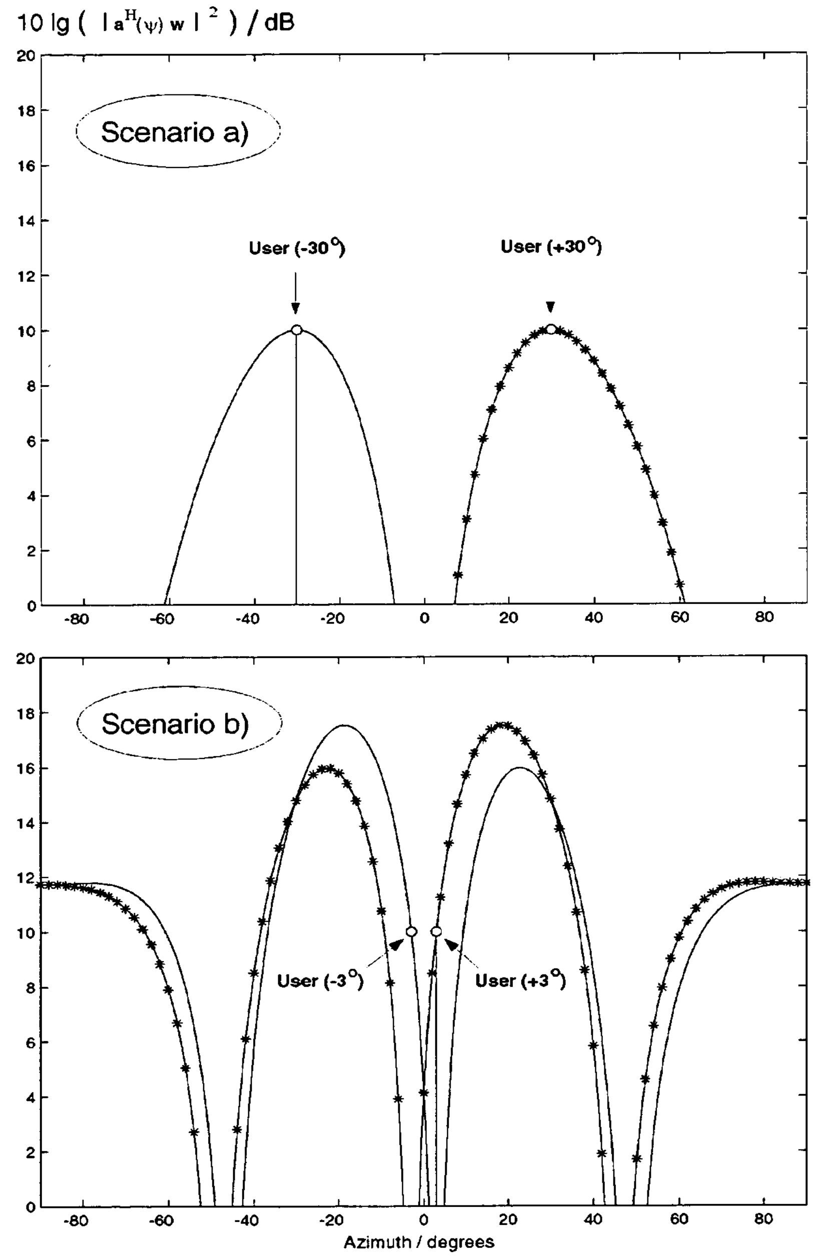


Figure 1. Beampatterns created by a 4-element ULA for a spatially well separable scenario (a) and a spatially badly separable scenario (b).

The downlink transmit power P is proportional to the squared length $w^H w$ of the weight vector applied at the array, whereas the RF receive power S at the mobile antenna is proportional to the term $w^H C w$. Therefore, the beamforming problem to calculate the optimum weight vector w can be mathematically put as the following constraint optimization problem:

$$\underset{w}{\text{minimize}} \left\{ P = w^H w \right\} \quad \text{subject to} \quad S = w^H C w = N \cdot SNIR. \tag{2}$$

The solution w of the above problem is proportional to the dominant eigenvector $\tilde{u}(C)$ of the spatial covariance matrix C, i.e. the eigenvector corresponding to the largest eigenvalue $\tilde{\lambda}(C)$ of C. Hence, the minimum downlink transmit power is given by the product $\tilde{\lambda}^{-1}(C) \cdot N \cdot SNIR$.

Generalizing this result to the case of K users (indexed by $(\cdot)_1 \cdots (\cdot)_K$) being accommodated in K separate channels leads to the minimum RF transmit power

$$P_{\min} = \sum_{k=1}^{K} \frac{N_k \cdot SNIR}{\tilde{\lambda}(C_k)}. \tag{3}$$

Let us now consider the SDMA case with K users operating in the same downlink channel. The beampatterns to separate the users $k = 1 \cdots K$ from each other on the downlink will be produced by the weights $w_1 \cdots w_K$. Generalizing (2), the corresponding SDMA beamforming problem turns out to be an optimization problem with K constraints which can be put as follows:

$$\underset{w_1 \ldots w_K}{\text{minimize}} \left\{ P = \sum_{k=1}^{K} w_k^H w_k \right\} \quad \text{subject to} \quad \frac{w_k^H C_k w_k}{SNIR} = N_k + \sum_{\substack{l=1 \\ l \neq k}}^{K} w_l^H C_k w_l \;\; \forall \; k = 1 \ldots K. \tag{4}$$

3.2. MEASURING THE SPATIAL SEPARABILITY

If $SNIR, N_1, \cdots, N_K > 0$ holds and all spatial covariance matrices C_k are non-negative definite (following from the definition (1)), the downlink transmit power P solving the SDMA beamforming problem (4) will never be lower than the minimum power $P_{\min}$ defined in (3).

The case $P = P_{\min}$ is characterized by the dominant eigenvector of each user's spatial covariance matrix being orthogonal to all other users' spatial covariance matrices:

$$C_k \, \tilde{u}(C_l) = 0 \;\; \forall \; k = 1 \ldots K, \; l \neq k = 1 \ldots K. \tag{5}$$

Plugging these conditions into (4) shows that in this case all weight vectors w_k will be proportional to the dominant eigenvectors $\tilde{u}(C_k)$, like in the single user case (2). The downlink beamformer only produces useful signal power but no cross interference, since $w_l^H C_k w_l = 0$ holds for all $k \neq l$. Therefore, it makes sense to refer to the case incorporated in (5) as "perfect spatial separability".

Simulations have shown that for "imperfect spatial separability" there is a connection between the angular distance of the users and the difference between the downlink transmit power P and the theoretical minimum $P_{\min}$. As an example, let us assume $K = 2$ users have to be separated on the SDMA downlink by beamforming with an $M = 8$ element ULA. The channels of both users can be described by means of a single DOA each. The corresponding attenuations ($\rho_1 = \rho_2 = 0$ dB) and the elevation angles ($\theta_1 = \theta_2 = 0^0$) are identical. Both DOAs are considered to be made up of very large numbers Q_1 and Q_2 of propagation pathes

with Gaussian distributed azimuths. The mean azimuths ψ_1 and ψ_2 are different, whereas the corresponding standard deviations ("angular spreads" [12]) are identical ($s_1 = s_2 = s$).

Figure 2 shows the ratio $P/P_{\min}$, with the first user constantly looking at the array boresight ($\psi_1 = 0^0$), and the azimuth ψ_2 of the second user varying from -90^0 to $+90^0$. Ignoring the sidelobes in Figure 2, the ratio $P/P_{\min}$ tends to increase, if the DOAs of the two users move closer to each other. Finally, if the DOAs are just too close (or even identical), the problem (4) does not yield any solution at all, i.e. the users are spatially no longer separable by beamforming.

Therefore, it makes sense to use the ratio $\eta = P/P_{\min}$, which from now on will be refered to as the "cross interference loss", as a measure for the spatial separability of a scenario. Doing this, $\eta = 0$ dB means optimum spatial separability, whereas $\eta \to \infty$ refers to the case of no spatial separability at all.

Note, that since spatial covariance matrices are the input parameters of the beamforming problem (4), the definition of the cross interference loss is not restricted to single DOA scenarios or discrete DOA scenarios. The cross interference loss can rather be calculated for any narrowband scenario and any antenna array configuration.

3.3. Calculating the Cross Interference Loss

The cross interference loss η depends on the downlink transmit power P and is, therefore, only known after the SDMA beamforming problem (4) has been solved. Unfortunately, this problem is numerically tricky, since both the target function and the constraints are non-linear functions of the variables $\boldsymbol{w}_1 \cdots \boldsymbol{w}_K$.

Computationally inexpensive algorithms yielding approximations of the solution of the SDMA beamforming problem have been presented in [13] and [14]. The computational costs of these algorithms are basically caused by K generalized (Hermitian) $M \times M$ eigenvalue decompositions (GEVDs). If a DOA sensitive channel allocation scheme has to estimate the cross interference loss η in each channel $l = 1 \cdots L$, altogether $L + K^{(1)} + \cdots + K^{(L)}$ GEVDs have to be executed for each user request. The resulting computational effort might be critical in real time, especially in large systems with a high number L of channels.

Therefore, we will present a new method to quickly calculate an approximation of the solution of (4). We will refer to it as the "eigenvector approximation", since the basic idea of this algorithm is to restrict the weight of each user in a way that it is proportional to the dominant eigenvector of the corresponding spatial covariance matrix:

$$\boldsymbol{w}_k = \sqrt{P_k}\,\tilde{\boldsymbol{u}}(\boldsymbol{C}_k) \;\; \forall \;\; k = 1 \ldots K. \tag{6}$$

This way the degrees of freedom are reduced to a great extent, since the complex-valued $M \times 1$ variables $\boldsymbol{w}_1 \cdots \boldsymbol{w}_K$ will be replaced by the real-valued 1×1 variables $P_1 \cdots P_K$.

Plugging the extra constraints (6) into (4) reveals that the originally non-linear constraint minimization problem melts down to an unconstraint linear problem:

$$\frac{P_k\,\tilde{\boldsymbol{u}}^H(\boldsymbol{C}_k)\,\boldsymbol{C}_k\tilde{\boldsymbol{u}}(\boldsymbol{C}_k)}{SNIR} = N_k + \sum_{\substack{l=1\\ l\neq k}}^{K} P_l\,\tilde{\boldsymbol{u}}^H(\boldsymbol{C}_l)\,\boldsymbol{C}_k\tilde{\boldsymbol{u}}(\boldsymbol{C}_l) \;\; \forall \;\; k = 1 \ldots K. \tag{7}$$

Solving (7) yields the variables $P_1 \cdots P_K$ and the approximated downlink transmit power $\tilde{P} = P_1 + \cdots + P_K$. Like this, the computational costs for each user request are basically resulting from the need to compute L dominant eigenvectors (one in each channel $l = 1 \cdots L$)

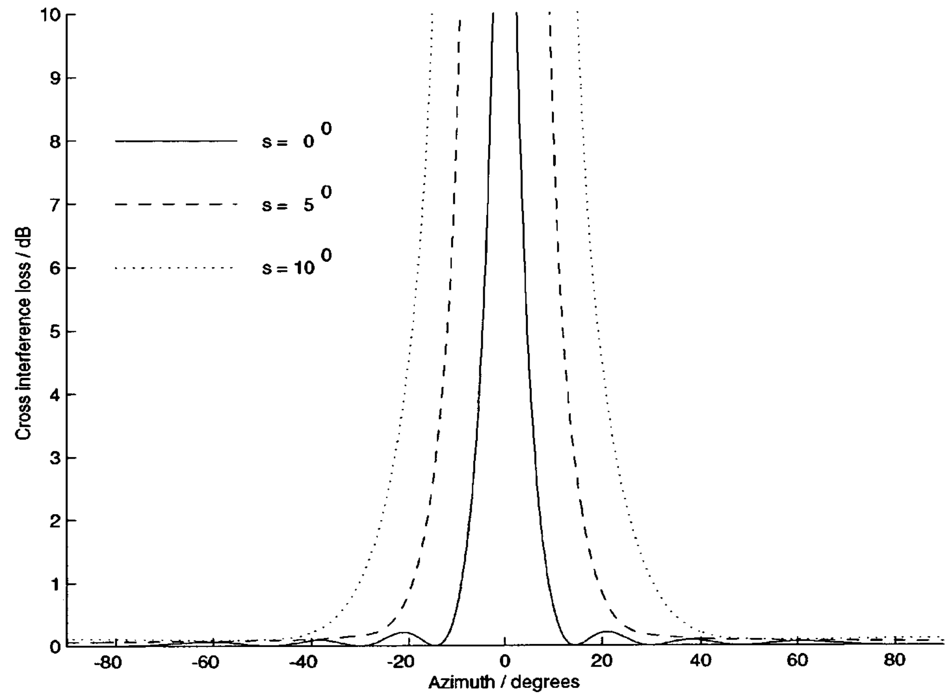


Figure 2. The cross interference loss η versus the azimuth ψ_2 of the second user.

for the matrices $C_k^{(l)}$ of the new user. Note that in mobile radio systems characterized by spatial covariance matrices that are almost identical for all channels (which is true for GSM and most of the other present systems), only one dominant eigenvector has to be calculated for each user request.

Figure 3 shows the approximated cross interference loss $\tilde{\eta} = \tilde{P}/P_{\min}$ for the scenario described in Section 3.2. Comparing Figures 3 and 2 exhibits that the eigenvector approximation works particularly well for spatially well separable scenarios, characterized by a small cross interference loss η. This is not surprising, since in the case of perfect spatial separability ($\eta = 0$ dB) the eigenvector approximation yields the exact solution to the original SDMA beamforming problem (see Section 3.2.

Evaluating Figure 3 for scenarios with high cross interference reveals that the eigenvector approximation produces more "pessimistic" results about the spatial separability of scenarios than an exact algorithm would do. But since in practice only spatially well separable scenarios with small cross interference losses η are relevant for downlink communication, the eigenvector approximation suggests to be an efficient tool to evaluate the spatial separability of a mobile radio scenario.

3.4. The Eigenvector Method

Finally, the *Eigenvector Method*, a channel allocation scheme based on the considerations in the previous sections, is given by the following procedure:

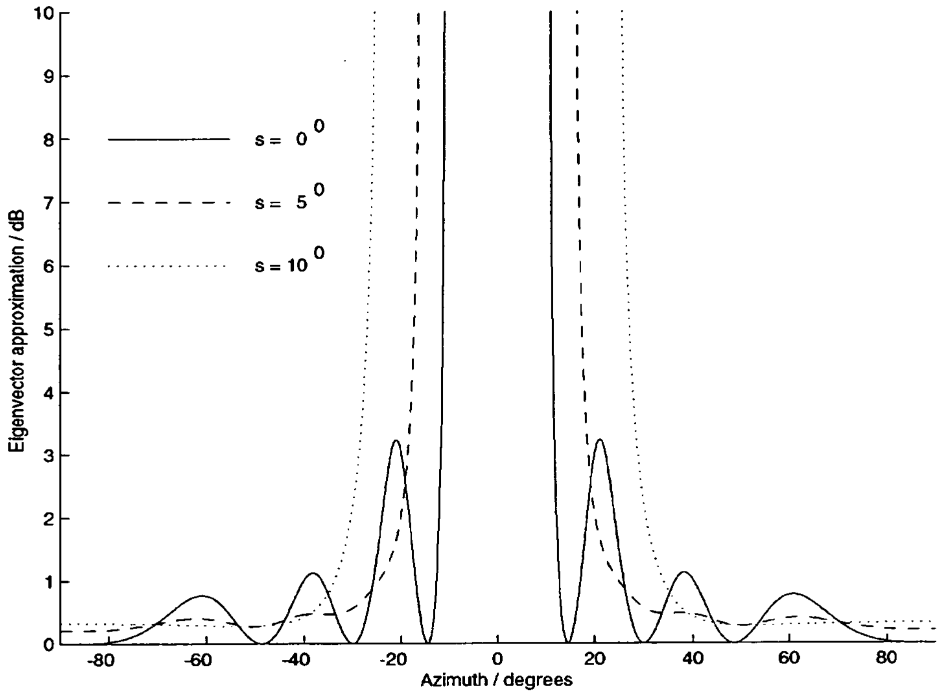


Figure 3. The eigenvector approximation $\tilde{\eta}$ of the cross interference loss versus the azimuth ψ_2 of the second user.

---

1. Choose appropriate values for the desired signal-to-noise-and-interference ratio *SNIR* and for the maximum tolerable cross interference loss $\tilde{\eta}_{\max}$.
2. Estimate the spatial covariance matrix C of the new user requesting for a communication channel.
3. Compute a dominant unit eigenvector $\tilde{u}(C)$ of C.
4. For all channels $l = 1 \cdots L$ do:

   - Estimate the receiver noise $N^{(l)}$ at the location of the new user's mobile in the specific channel l. In general, this value $N^{(l)}$ predominantly results from CCI from neighboring cells and has to be measured by the mobile and communicated to the base station.
   - Assume the new user will be allocated to the channel l, so that the channel will have $K = K^{(l)} + 1$ users indexed by $(\cdot)_1 \cdots (\cdot)_K$. With all spatial covariance matrices $C_1^{(l)} \cdots C_K^{(l)}$, the corresponding dominant unit eigenvectors $\tilde{u}(C_1^{(l)}) \cdots \tilde{u}(C_K^{(l)})$ and the noise powers $N_1^{(l)} \cdots N_K^{(l)}$ known to the base station, solve the real-valued $K \times K$ system

$$\begin{pmatrix} \frac{\tilde{u}^H(C_1^{(l)})\, C_1^{(l)}\, \tilde{u}(C_1^{(l)})}{N_1^{(l)}\, SNIR} & \cdots & -\frac{\tilde{u}^H(C_K^{(l)})\, C_1^{(l)}\, \tilde{u}(C_K^{(l)})}{N_1^{(l)}} \\ \vdots & \ddots & \vdots \\ -\frac{\tilde{u}^H(C_1^{(l)})\, C_K^{(l)}\, \tilde{u}(C_1^{(l)})}{N_K^{(l)}} & \cdots & \frac{\tilde{u}^H(C_K^{(l)})\, C_K^{(l)}\, \tilde{u}(C_K^{(l)})}{N_K^{(l)}\, SNIR} \end{pmatrix} \begin{pmatrix} P_1^{(l)} \\ \vdots \\ P_K^{(l)} \end{pmatrix} = \begin{pmatrix} 1 \\ \vdots \\ 1 \end{pmatrix}. \qquad (8)$$

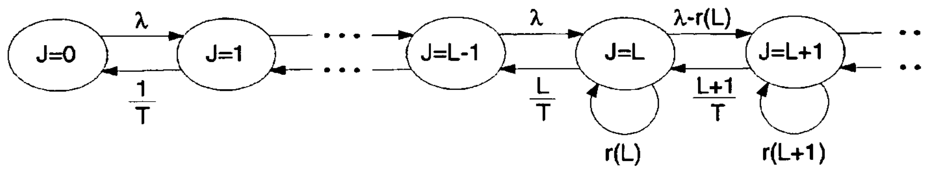


Figure 4. Traffic model for an SDMA system with L channels.

- If the system does not have a (unique) solution or any entry $P_k^{(l)}$ is non-positive, the K users in the channel l will be considered spatially unseparable. Otherwise, calculate the eigenvector approximation $\tilde{\eta}^{(l)}$ of the cross interference loss:

$$\tilde{\eta}^{(l)} = \frac{\sum_{k=1}^{K} P_k^{(l)}}{SNIR \sum_{k=1}^{K} \tilde{\lambda}^{-1}(C_k^{(l)}) \cdot N_k^{(l)}}. \tag{9}$$

5. Select the optimum channel $l_{\text{opt}} \in \{ 1 \cdots L \}$ according to the smallest approximated cross interference loss.
6. If $\tilde{\eta}^{(l_{\text{opt}})}$ is larger than the threshold $\tilde{\eta}_{\max}$, then
   reject the new user (since the system is supposed to be fully loaded),
   otherwise
   allocate him to the channel l_{opt} and calculate the corresponding beamforming weights:

$$w_k^{l_{\text{opt}}} = \sqrt{P_k^{(l_{\text{opt}})}}\, \tilde{u}(C_k^{(l_{\text{opt}})}). \tag{10}$$

The parameters $SNIR$ and $\tilde{\eta}_{\max}$ must be both chosen according to the system parameters and the type of radio channel. We assume that for present mobile radio air interfaces and Rayleigh fading channels an $SNIR$ between 10 and 20 dB and an $\tilde{\eta}_{\max}$ between 1 and 5 dB will guarantee robust downlink communication for every user in every channel.

4. Capacity

We will define the *capacity* of a mobile radio system in a traffic theory like manner:

> Capacity is the traffic $A = \lambda T$ an SDMA system can support without exceeeding a maximum blocking probability B during the channel allocation procedure.

In this context the calling rate (in calls per second) is denoted by λ, whereas T designates the average duration of a call (in seconds).

The corresponding traffic model is depicted in Figure 4: Each state in the Markov chain is characterized by the number J of users in the system. The transition from a state $J = j$ to a lower state $J = j - 1$ results from the termination of a call by one of the J users. Hence, the corresponding transition rate is given by the ratio j/T of the number of active users to the average call duration. The transition from a state $J = j$ to a higher state $J = j + 1$ is triggered by a user request resulting in a successful channel allocation. As long as there are still free channels available ($j < L$), the corresponding rate is identical to the calling rate

λ. If all channels are occupied with at least one user ($j \geq L$), there is a chance the channel allocator rejects the new user, represented by the rejection rate $r(j)$.

For $M > 1$, there is no way of describing the rejection rates $r(j)$ and the resulting overall blocking probability B by means of analytic formulae. Therefore, we had to resort to simulations in realistic SDMA scenarios to estimate B as a function of the traffic $A = \lambda T$.

5. Simulation Results

Our SDMA capacity predictions were based on a Monte Carlo simulation of a single mobile radio cell. The statistics of the parameters defining the radio channels between the users and the base station were chosen in compliance with the results of the 2D channel measurements carried out in the city of Munich in 1995 [15].

We are assuming a ring-shaped cell which has an SDMA base station in its center equipped with a uniform linear M element antenna array. The maximum distance from any user to the base station is given by the outer ring radius 5 km, the minimum distance by the inner ring radius 0.1 km. The user locations are uniformly distributed in the ring and independent from each other. Assuming a typical urban area, the average attenuation corresponding to the distance r_k of the user k to the base station can be approximated by $\overline{\rho}_k = 40\,lg(r_k/5m)$ dB. The corresponding attenuations are log-normally distributed with the Suzuki parameter $S = 6$ dB and the average $\overline{\rho}_k$.

The propagation paths of each user k are assumed to consist of either one group ($G_k = 1$, e.g. LOS[5] or one dominant reflector) or two groups ($G_k = 2$, e.g. LOS plus one dominant reflector or two dominant reflectors). The probabilities for both cases $G_k = 1$ and $G_k = 2$ are 0.5 each. Each group $g = 1 \cdots G_k$ is assumed to consist of $R_{kg} \in \{1 \cdots 50\}$ propagation paths.

The azimuths ψ_{kgr} and elevations θ_{kgr} of all paths $r = 1 \cdots R_{kg}$ are both Gauss-distributed with the means $\overline{\psi}_{kg}$ and $\overline{\theta}_{kg}$ and the standard deviations $s_\psi = s_\theta = 5^0$. The mean azimuths $\overline{\psi}_{kg}$ are uniformly distributed in the range $[-180^0; +180^0]$, whereas the mean elevations are constant ($\overline{\theta}_{kg} = 0^0$).

The noise powers $N_k^{(l)}$ are assumed to be made up by CCI from log-normally shadowed base stations 15–25 km away from the users. This interference model corresponds to a hexagonal non-SDMA cellular mobile radio network with a frequency reuse factor $r = 4$. The *Eigenvector method* operates with the signal-to-noise-and-interference ratio $SNIR = 10$ dB and the threshold $\eta_{\max} = 3$ dB.

The dependence of the blocking probability B on the traffic $A = \lambda T$ is depicted in Figure 5, with a constant number L of channels (top plot) and a constant number M of antennas (bottom plot).

Figure 6 shows the number L of separate channels an SDMA base station needs to support a given traffic A in the cell. The number of antennas was varied from $M = 1$ to $M = 16$. The number of calls simulated for each point in the plot is 10000. The tolerable blocking probability was $B = 1\%$.

Not surprisingly, Figure 6 can be interpreted in a way, that the higher the number M of antennas, the lower the number L of channels necessary to support a given traffic A. As an example, consider the traffic $A = 30$ erl: A conventional system ($M = 1$) like GSM needs

[5] Line of sight propagation.

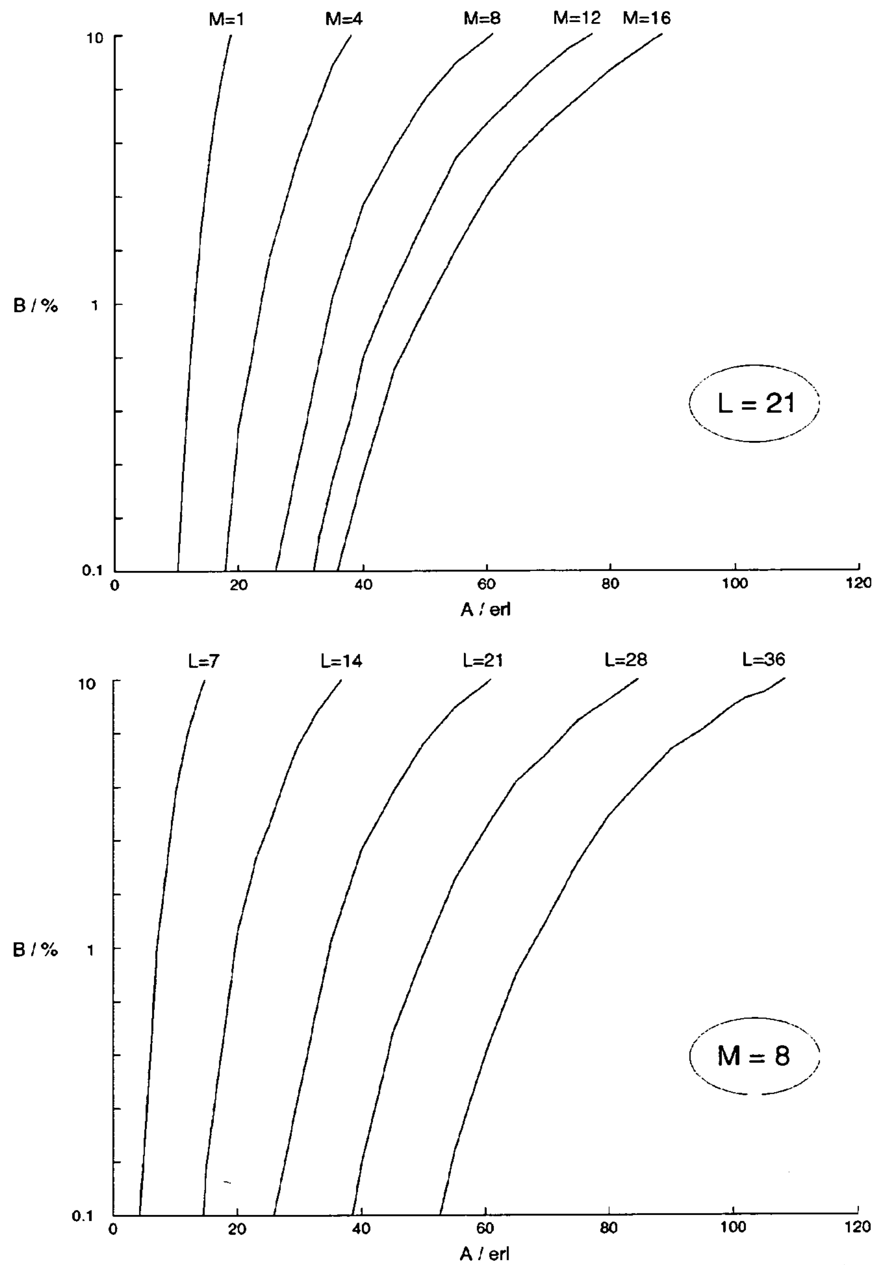


Figure 5. Blocking probability B versus traffic A.

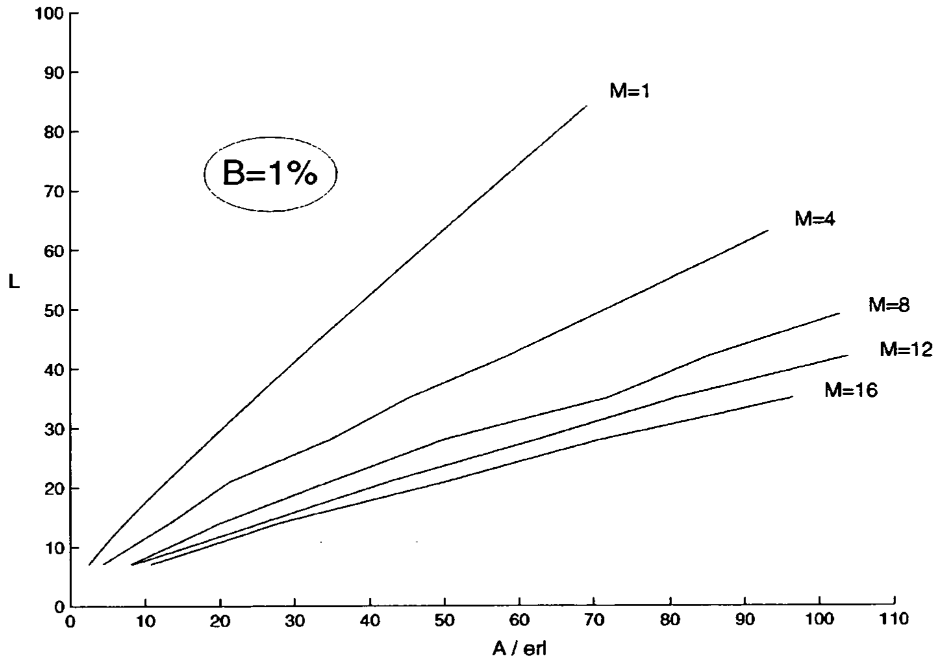


Figure 6. Supportable traffic A versus the number L of channels with a maximum blocking probability $B = 1\%$.

$L_1 = 42$ channels in order not to exceed the blocking probability $B = 1\%$. An SDMA system with $M = 8$ antennas needs $L_8 = 26$ channels and one with $M = 16$ antennas needs $L_{16} = 14$ channels. The corresponding results for $A = 60$ erl are: $L_1 = 73$, $L_8 = 32$ and $L_{16} = 24$.

6. Conclusions

In this paper, SDMA capacity is defined as the traffic A a system can support without exceeding a maximum blocking probability B during the channel allocation procedure. The capacity was estimated by simulating the performance of a specific SDMA channel allocation scheme, the *Eigenvector Method*, in a realistic urban mobile radio cell. We assumed that channel allocation and, hence, system capacity is limited by the maximum number of users which can be accommodated on the downlink in a robust way.

As a result, the simulations yielded the number L of channels (i.e. FDMA, TDMA or CDMA slots) a base station must provide to be capable of handling a given traffic A without exceeding the blocking probability $B = 1\%$ (see Figure 6). The capacity increase L_1/L_M over a conventional system with a single antenna is rather dependent on the number M of antennas than on the traffic A, as shown in Table 1 which has been extracted from the plot shown in Figure 6.

The SDMA capacity increase L_1/L_M refers to the RWC gain from managing more than one user in each channel of a mobile radio cell. Note that employing SDMA features in all cells of a mobile radio network will result in an even higher increase in total system capacity. The reason is that downlink beamforming with antenna arrays in each cell produces less CCI in

Table 1. The SDMA capacity increase L_1/L_M over a conventional system depending on the traffic A and the number M of antennas.

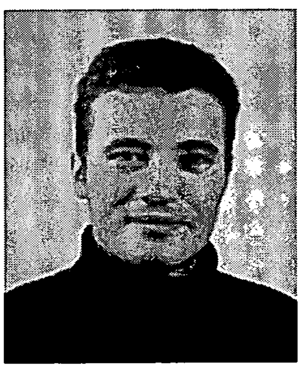
L_1/L_M	$A = 20$	$A = 40$	$A = 60$
$M = 1$	1.00	1.00	1.00
$M = 4$	1.49	1.66	1.73
$M = 8$	2.12	2.24	2.38
$M = 12$	2.53	2.63	2.74
$M = 16$	2.75	2.95	3.06

neighboring cells than omnidirectional transmission with single antennas. Therefore, system capacity can be further increased by reducing the frequency reuse factor employed in the network.

References

1. P. Jung, J. Blanz and P.W. Baier, "Coherent Receiver Antenna Diversity for CDMA Mobile Radio Systems Using Joint Detection", *The 4th Int. Symp. on Personal, Indoor and Mobile Radio Comm. (PIMRC 93)*, Yokohama, Japan, 1993.
2. X. Lasne, C. Baroux and G.K. Kaleh, *Joint Reception of Multi-User Data for Synchronous CDMA*, COST231(92)85, 1992.
3. C. Farsakh and J.A. Nossek, "Application of SDMA to Mobile Radio", *The 5th Int. Symp. on Personal, Indoor and Mobile Radio Comm. (PIMRC 94)*, The Hague, The Netherlands, pp. 736–739, 1994.
4. C. Farsakh and J.A. Nossek, "Data Detection and Channel Allocation on the Uplink of an SDMA Mobile Radio System", *Vehicular Technology Conference (VTC 97)*, Phoenix AZ, U.S.A., 1997.
5. C. Brunner, M. Haardt, C. Farsakh and J.A. Nossek, "A Two Step Spatio-Temporal Filtering Scheme for Smart Antennas Compared to a Multi-User Maximum Likelihood Algorithm", submitted to *The IEEE 48th Vehicular Technology Conference (VTC 98)*, Ottawa, Canada, 1998.
6. H.P. Lin, S.S. Jeng, I. Parra, G. Xu, W.J. Vogel and G.W. Torrence, "Experimental Studies of SDMA Schemes for Wireless Communication", *The 1995 International Conference on Acoustics, Speech and Signal Processing (ICASSP 95)*, IEEE Press, Detroit, pp. 1760–1763, 1995.
7. S.S. Jeng, H.P. Lin, G. Xu and W.J. Vogel, "Measurements of Spatial Signatures of an Antenna Array", *The 6th Int. Symp. on Personal, Indoor and Mobile Radio Communications (PIMRC 95)*, Toronto, Canada, pp. 669–672, 1995.
8. J. Wigard, P.E. Mogensen, F. Frederiksen and O. Nørklit, "Evaluation of Optimum Diversity Combining in DECT", *The 6th Int. Symp. on Personal, Indoor and Mobile Radio Communications (PIMRC 95)*, Toronto, Canada, pp. 507–511, 1995.
9. M. Haardt and J.A. Nossek, "Unitary ESPRIT: How to Obtain Increased Estimation Accuracy with a Reduced Computational Burden", *IEEE Trans. Signal Processing*, Vol. 43, pp. 1232–1242, 1995.
10. M. Haardt, M.D. Zoltowski, C.P. Mathews and J.A. Nossek, "2D Unitary ESPRIT for Efficient 2D Parameter Estimation", *Proc. Int. Conf. Acoust., Speech, Signal Processing (ICASSP 95)*, Detroit MI, U.S.A., pp. 2096–2099, 1995.
11. C. Farsakh and J.A. Nossek, "Maximizing the Capacity of an SDMA Mobile Radio System", *Int. Conf. Telecommunication (ICT 97)*, Melbourne, Australia, 1997.
12. P.C. Eggers, "Angular Dispersive Mobile Radio Environments Sensed by Highly Directive Antennas", *The 6th Int. Symp. on Personal, Indoor and Mobile Radio Comm. (PIMRC 95)*, Toronto, Canada, pp. 522–526, 1995.

13. P. Zetterberg, "A Comparison of Two Systems for Downlink Communication with Antenna Arrays at the Base", *IEEE Wireless Communications Systems Symposium 1995*, Long Island, New York, 1995.
14. C. Farsakh and J.A. Nossek, "Channel Allocation and Downlink Beamforming in an SDMA Mobile Radio System", *The 6th Int. Symp. on Personal, Indoor and Mobile Radio Comm. (PIMRC 95)*, Toronto, Canada, pp. 687–691, 1995.
15. A. Klein, W. Mohr, R. Thomas, P. Weber and B. Wirth, "Directions-of-Arrival of Partial Waves in Wideband Mobile Radio Channels for Intelligent Antenna Concepts", *The IEEE 46th Vehicular Technology Conference (VTC 96)*, Atlanta GA, U.S.A., pp. 849–853, 1996.



Christof Farsakh received his diploma and Ph.D. degree in electrical engineering from the Technical University of Munich, Germany, in 1992, and 1997, respectively. His research interests were signal processing, non-linear optimization and the application of smart antenna technologies in mobile radio.

He is now working for an intellectual property firm in Munich, specializing in patents for inventions supporting new technologies for future mobile radio systems.

He was the co-winner of the 1990 innovation prize of the "Mannesmann-Mobilfunkstiftung" foundation in recognition of his work on smart antennas in mobile radio communications.

Josef A. Nossek, born in Vienna, Austria, on 17 December 1947 (S'72 – M'74 – SM'81 – F'93) received the Dipl.-Ing. and Dr. degrees, both in electrical engineering, from the Technical University of Vienna, Austria, in 1974 and 1980, respectively.

In 1974 he joined SIEMENS AG, Munich, Germany, where he was engaged in the design of passive and active filters for communication systems. In 1978 he became a supervisor, in 1980 head of a group of laboratories concerned with the design of monolithic filters (analog and digital) and electromechanical and microwave filters. Since 1982 he has been head of a group of laboratories designing digital radio systems within the Transmission Systems Department. In 1984 he spent a month as a Visiting Professor at the University of Capetown.

From 1987–1989 he was head of the Radio Systems Design Department, where he was instrumental in introducing high speed VLSI signal processing into digital microwave radio. Since April 1989 he is Professor for Circuit Theory and Design at the Technical University of Munich. He is teaching undergraduate and graduate courses in the field of circuit and system theory and conducting research in the areas of real-time signal processing, neural networks and dedicated VLSI-architectures.

He has published more than 100 papers in scientific and technical journals and conference proceedings. He holds a number of patents. In 1988 he received the ITG prize and was the co-winner of the 1998 innovation prize of the "Mannesmann-Mobilfunkstiftung" foundation in recognition of his work on smart antennas in mobile radio communications. He has been a member of numerous organizing and program committees and is member of editorial boards of several scientific journals.